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INVESTIGATION OF A NEW CLASS OF LOW-PROFILE

MULTI-LAYER PRINTED ANTENNAS

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INVESTIGATION OF A NEW CLASS OF LOW-PROFILE MULTI-LAYER PRINTED ANTENNAS

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TABLE OF CONTENTS

I. RESEARCH HIGHLIGHTS, 1997-2001

Over the last four years, our research work funded by this grant has resulted in numerous technical contributions that have been detailed in four annual technical reports, submitted to ONR, as well as in 20 journal and symposium publications. Highlights of our contributions are as follows:

1) Development of novel gain enhancement methods for multi-layered printed antennas.
We have developed two techniques to achieve high gain operation for printed antennas. The first technique uses a gain enhancement method based on the resonance condition of a single microstrip patch in a multilayer geometry with a high permittivity focusing layer, whereas the second alternative technique uses a Yagi-like configuration of the stacked patches printed in a low permittivity multilayer medium. These schemes are particularly well-suited for millimeter-wave applications because of their inherent low conductor losses. The application of these methods in design of microstrip patch antennas with various feeding techniques such as probe-fed, proximity-fed and aperture-fed have been fully investigated. In addition, the feasibility of obtaining circular and dual polarization in conjunction with these gain-enhancement methods have been demonstrated. Furthermore, the radiation performance and the effect of mutual couplings and surface-waves, when these printed antennas are used as elements in an array environment, have been analyzed. Finally, we have also extended these high gain methods to antennas printed in a multi-layer anisotropic medium, where it is shown that by a proper selection of anisotrpy ratios and thickness of the layers, it is possible to obtain two separate high gain beams in E and H planes.

2) Development of an electromagnetic optimization engine for multi-layer printed patch antennas of arbitrary shapes.
This numerical engine is based on a hybrid combination of Method of Moments (MOM) and Evolutionary Programming (EP). A novel feature of the MOM portion of this numerical engine is a new semi-analytical technique that speeds up the evaluation of the corresponding Green's functions by a factor of 10 or higher. For the optimization part of the engine, we have developed various EP algorithms for continuous, discrete and mixed parameter optimizations. Also, we have developed efficient methods which combine two different mutation operators, namely

Gaussian and Cauchy, in the EP optimization process. The developed electromagnetic optimization engine has been instrumental in design of the gain-enhanced multi-layered antennas described above.

3) Analysis of the curvature effects on performance of the gain-enhanced printed antennas.
The parameters that influence the gain enhancement of a cylindrical microstrip patch antenna covered with a multi-layered dielectric medium have been studied in detail by developing an analytically modeling each edge of the patch by a magnetic line source. Variation in directivity and normalized radiation conductance as a function of curvature radius and thickness of layers has been investigated for different values of dielectric constants. It is shown that curvature effect could significantly change the maximum gain conditions for a planar multi-layered microstrip antenna. We are also currently in the midst of developing a numerical code, based on the finite-difference time-domain (FDTD) method, for accurate characterization of the multi-layer microstrip antennas on cylindrical surfaces, including the effects of truncated dielectric layers. Upon completion, this code can be used for radiation as well as RCS analysis of the Yagi-like antennas printed in a curved multi-layered dielectric medium.

4) Development of antenna measurement facility at Villanova.
The measurement facility of the Antenna Research Laboratory (ARL), which has been partially funded by the present grant, has now been completed. This state-of-the art indoor facility includes a 29' x 27' x 28' anechoic chamber (compact range) capable of fully automated high quality pattern and gain measurements from 1-40 GHz. We have teamed with Naval Sea Systems Command, NAVSEA-Philadelphia in a joint effort to extend the capabilities of this facility to 75 GHz and include accurate radar cross section (RCS) measurements. This Academia/Government relationship allows Villanova's ARL to offer extensive facilities and expertise to Government and Industry for the development, integration, evaluation and testing of RCS and antenna components and systems.

Because the annual reports submitted to the ONR in 1998, 1999 and 2000 have detailed most of our research work in this project, the present report only covers the progress made during the October 2000 - September 20001 time period.

A. Publications based on the present research grant

1.   A. Hoorfar, "Analysis of a 'Yagi-like' printed stacked dipole array for high gain applications," *Microwave and Optical Technology Letters*, Vol 17, No. 5, pp. 317-321, April 1998.

2.   A. Hoorfar, "A 'Yagi-like' gain enhancement method for multi-layer printed antennas," Proceedings of the *1998 International Symposium on Electromagnetic Theory* , PP. 250-252, May 1998, Thessaloniki, Greece.

3.   A. Hoorfar and D. C. Chang,"Semi-analytical solutions for microstrip Green's functions in multi-layered media," Proceedings of the *1998 International Symposium Electromagnetic Theory* , PP. 618-620, May 1998, Thessaloniki, Greece.

4.   K. Chellapilla and A. Hoorfar, "Evolutionary programming: an efficient alternative to genetic algorithms for electromagnetic optimization problems," Proceedings of the *IEEE AP-S International Symposium*, Atlanta, GA, pp. 42-45 , June 1998.

5.   A. Hoorfar and K. Chellapilla, "Gain optimization of a multi-layer printed dipole array using evolutionary programming," Proceedings of the *IEEE AP-S International Symposium*, Atlanta, GA, pp. 46-49 , June 1998.

6.   A. Hoorfar, "Integral equation modeling of microstrip circuits and antennas using closed-form Green's Functions," *Microwave and Optical Technology Letters*, Vol. 20, Number 5, pp.342-345, March 1999.

7.   A. Hoorfar, G. Girard and A. Perrotta, "A dual frequency circular-polarized electromagnetically-fed microstrip antenna,' *Electronics Letters*. Vol. 35, Number 10, pp.759-761, May 1999.

8.   A. Hoorfar, "Mutation-based evolutionary algorithms and their applications to optimization of antennas in layered media," Proceedings of the *IEEE AP-S International Symposium*, Orlando, FL, pp. 2876-2879, July 1999.

9.   R. G. Holland and A. Hoorfar, "Gain enhancement method for microstrip antennas in uniaxial anisotropic media," *Digest of the 26th General Assembly of the International Union of Radio Science (URSI)*, Toronto, Canada, August 1999.

10.   K. Chellapilla and A. Hoorfar, "Intelligent operator design in evolutionary computation and its application to antenna optimization problems," *Digest of the 26th General Assembly of the International Union of Radio Science (URSI)*, Toronto, Canada, August 1999.

11.   A. Hoorfar and Yuan Liu, "A study of Cauchy and Gaussian mutation operators in evolutionary programming optimization of antenna structures," Proceedings of the *16th Annual Applied Computational Electromagnetics Conference* , Monterey, CA, pp. 63-69, March 2000.

12   Yuan Liu and A. Hoorfar, "Optimization of antenna arrays using evolutionary programming", Proceedings of the *2000 IEEE Sarnoff Symposium*, Ewing, NJ, pp. 147-150, March 2000.

13   C. Wan and A. Hoorfar, "Improved Design Equations for Multilayer Microstrip Lines," *IEEE Microwave and Guided Wave Letters*,, Vol. 10, Number 6, pp. 223-24, June 2000..

14   A. Hoorfar and Yuan Liu, "Antenna Optimization Using an Evolutionary Programming Algorithm with a Hybrid Mutation Operator," Proceedings of the *IEEE AP-S International Symposium*, Salt Lake City, UT, pp. 1026-1029, July 2000.

15   A. Hoorfar and A. Perrotta, "An Experimental Study of Microstrip Antennas on Very High Permittivity Substrate and Very Small Ground-Planes, " *IEEE Transactions on Antennas and Propagation* , pp. 838-840, April 2001.

16   A. Hoorfar and Yuan Liu, "Evolutionary design of a Yagi-like array of proximity-fed stacked microstrip patch elements," Proceedings of the *URSI International Symposium on Electromagnetic Theory*, Victoria, Canada, pp. 630-632, May, 2001.

17   A. Hoorfar, S. Nelaturi and J. Zhu, "Electromagnetics optimization using an evolutionary algorithm with a mixed-parameter self-adaptive mutation operator," Proceedings of *the IEEE AP-S International Symposium*, Boston, MA, pp. 634-637, July 2001.

18   A. Hoorfar and R. Sun, "High gain microstrip antennas with a high permittivity focusing layer for millimeter-wave applications," Proceedings of the *IEEE International Conference on Electromagnetics in Advanced Applications*, Torino, Italy, pp. 255-258, September, 2001.

19   S. Nelaturi and A. Hoorfar, "Closed form expressions for Green's functions of microstrip structures on a uniaxially anisotropic substrate," submitted to *IEEE Transactions on MTT*, June 2001.

20   A. Hoorfar, K. Challapilla and S. S. Rao,"Evolutionary Programming: an alternative to Genetic Algorithms in electromagnetic optimization problems," To be submitted to *IEEE Transactions on antennas and Propagation.*

II. SUMMARY (2000-2001)

This progress report summarizes our research efforts on modeling, analyses and optimization of multi-layered printed antennas for various applications in communications and radar that may require high gain, wide bandwidth and/or multi-band operations. The report covers the time-period, October 2000 – September 2001. The technical progress made during this period include:

i) A gain enhancement method under investigation in this project has been based on the resonance condition of a single microstrip patch in a multilayer geometry with a high permittivity focusing superstrate layer.. We have now extended this high gain technique to design of proximity-fed and aperture-fed antennas for millimeter-wave applications like local multi-point distribution system (LMDS) and hybrid fiber-optic wireless system. Design optimization, based on method of moments, for the case of a four-layer proximity-coupled patch at a frequency of 31 GHz is presented and its performance in terms of gain, pattern and surface wave efficiency is analyzed. It is shown that conductor losses, using this gain enhancement technique, is significantly less than that in a standard series or corporate fed microstrip array with a similar gain. The mutual coupling effects in a two- element array of such radiating structures are also studied, where it is found that the presence of the high permittivity layer significantly increases the coupling between the elements. In addition, an alternative high gain technique, based on the Yagi-like concept, that utilizes parasitic elements at the interface between layers, and provides high gain without the use of high permittivity layers, has also been investigated in detail for millimeter-wave applications.

ii) Effect of curvature on gain enhancement of a multi-layered microstrip antenna is investigated by analyzing the radiated fields of a rectangular patch in a multi-layered cylindrical surface. Variation in directivity and normalized radiation conductance as a function of curvature radius and thickness of layers has been investigated for different values of dielectric constants. It is shown that curvature effect could significantly change the maximum gain conditions previously reported in the literature for a planar multi-layered microstrip antenna. . Development of a FDTD code for accurate characterization of the multi-layer microstrip antennas on cylindrical surfaces, including the effects of truncated dielectric layers, is presently in progress.
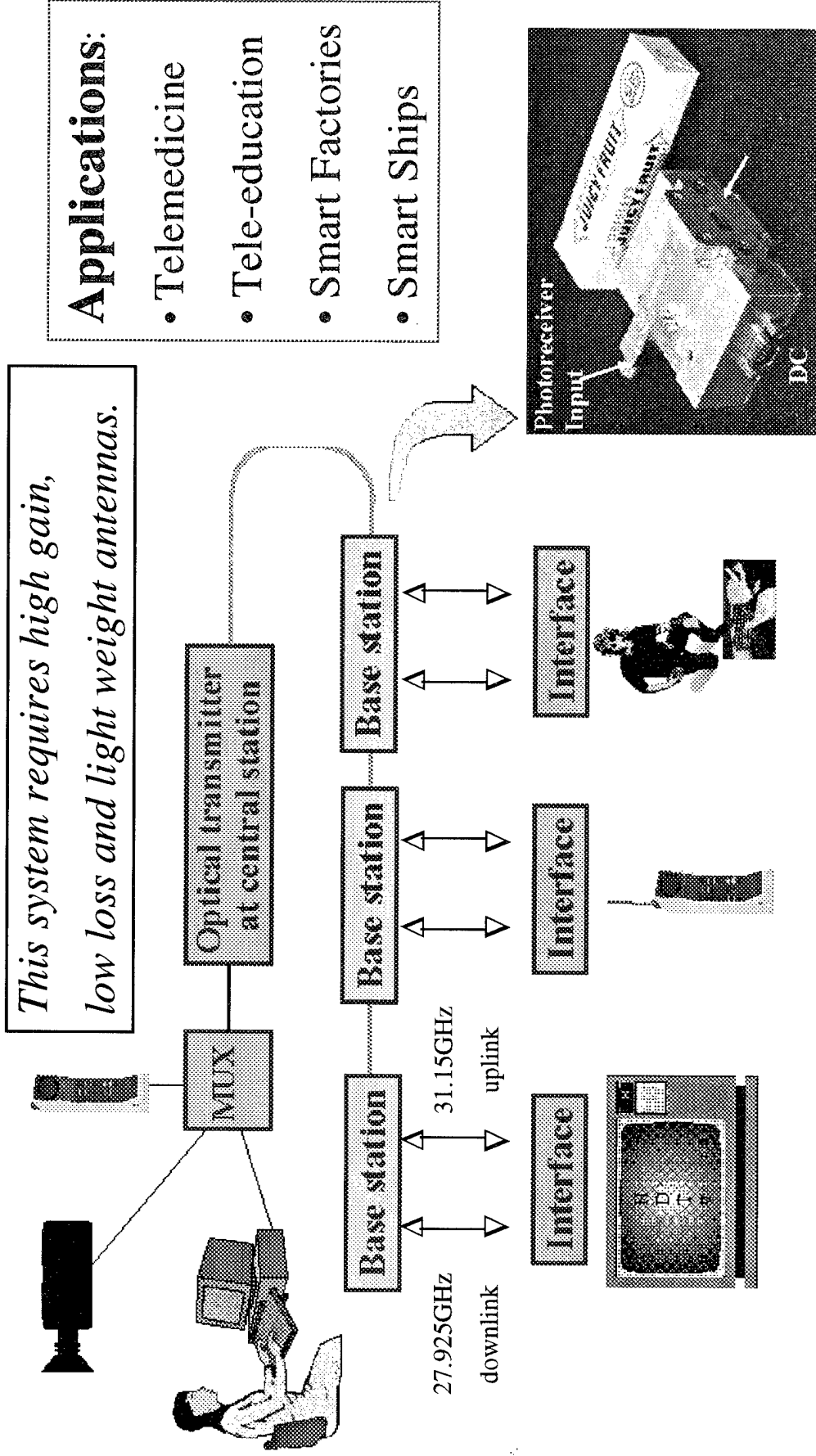
III. HIGH GAIN MULTILAYERED MICROSTRIP PATCH ANTENNAS FOR MILLIMETERWAVE APPLICATIONS

The broadband information applications including high speed internet services, wireless cable TV / modem, tele-medicine and digital battlefield are in the midst of pushing the wireless systems to the upper microwave and into the millimeter-wave frequency regions. Technologies like local multi-point distribution system (LMDS) and hybrid fiber-optic wireless system [1] are being developed at millimeter-wave frequencies for commercial as well as secure military applications. These systems would require high gain, low loss, low profile and lightweight antennas to be deployed in pico-cell base stations installed in outdoor or indoor environments. In indoor applications the base stations would be located in ceilings of offices and manufacturing facilities. For common transmit-receive operation or to combat multipath these systems may also require dual frequency and/or dual polarization for diversity. These requirements provide challenging problems in design of the future generation of communication antennas.

Figure 1 shows the system diagram of the system proposed in [1] with receive and transmit frequencies of 27.925 GHz and 31.15 GHz, respectively. The base-station and the trans-receiver are both very small and require compact printed antennas with a gain of 13 dBi or higher. Conventional array methods employing corporate or series feeding techniques have been widely used for design of high gain printed antennas in millimeter-wave frequencies. Printed arrays, however, suffer from high conductor as well as surface-wave losses that could result in very low antenna efficiency. In addition, these arrays may result in antennas occupying an electrically large real estate, which may not be desirable for some applications. An alternative technique for gain enhancement is based on the resonance condition of a single microstrip radiator in multi-layer geometry with alternating low and high permittivity dielectric layers [2,3].

We previously investigated the feasibility of the method in designs of rectangular rmicrostrip with various feeding techniques, including, coaxial, proximity and aperture feedings for lower microwave frequency applications [4]. In addition, application of these schemes to design of high gain antennas with circular and dual polarization was researched. The total thickness of the resulting structures, however, are impractical for lower microwave frequency

9

Hybrid Fiber Optic/Wireless Communication System

Applications:
- Telemedicine
- Tele-education
- Smart Factories
- Smart Ships

This system requires high gain, low loss and light weight antennas.

MUX

Optical transmitter at central station

Base station

Base station

Base station

Interface

Interface

Interface

Photoreceiver input

31.15GHz uplink

27.925GHz downlink

ranges. Extension of the concepts into millimeter-wave region, however, should result in low profile high gain radiating elements that would be attractive for systems such as the one in Figure 1. In this work we have investigated feasibility of proximity-coupled and aperture feeding schemes in conjunction with this gain enhancement method for millimeter-wave applications. In particular, design optimization of a four-layer proximity-fed microstrip structure is presented, and behaviors of its gain, radiation efficiency and surface wave power are analyzed. In addition, the mutual coupling effects between two elements of such high gain structures are investigated. Finally, a technique, which combines this gain enhancement method with a Yagi-like effect by utilizing parasitic elements at the interface between layers, is presented.

A High Gain Conditions for a Printed Dipole

There are two types of high gain conditions for a Hertzian electric dipole embedded in a multi-layered magneto-dielectric medium, shown in Figure 2 [2,3].
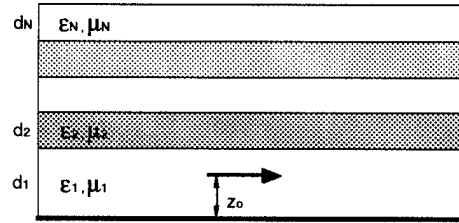
Fig. 2: An embedded dipole in a multi-layered medium.

Type I condition requires that odd layers have high wave-impedance (i.e., large μ_r) , even layers have low-wave impedance (i.e., large ε_r) and,

$$\frac{n_1 d_1}{\lambda_0} = \frac{1}{2} \quad , \quad \frac{n_1 z_0}{\lambda_0} = \frac{1}{4} \quad ; \quad \frac{n_i d_i}{\lambda_0} = \frac{1}{4} \quad , i \geq 2 \tag{1}$$

Type II condition requires that odd layers have low wave-impedance (i.e., large ε_r) , even layers have high wave-impedance (i.e., large μ_r) and,

$$\frac{n_1 d_1}{\lambda_o} = \frac{1}{4} \quad , \quad \frac{n_1 z_0}{\lambda_o} = \frac{1}{4} \quad ; \quad \frac{n_i d_i}{\lambda_0} = \frac{1}{4} \quad , \quad i \geq 2 \tag{2}$$

where $n_i = \sqrt{\varepsilon_{ri}\,\mu_{ri}}$ is the refractive index of layer i. It is obvious that for a purely dielectric medium (i.e. $\mu_r=1$), the dielectric layers should alternate between low and high permittivity layers.

The conditions (1) or (2) require a thick substrate thickness z_0, which is not desirable in a practical feeding scheme as it may result in spurious radiation from excitation probe or microstrip feed-line. It is, however, still possible to obtain high gain by optimizing the thickness of the dielectric layers around the values suggested in (1)-(2). Figure 3 shows the gain for a three-layer configuration with a finite length center-fed microstrip dipole located on a thin substrate with a thickness of $z_0 = 0.254$ mm and $\varepsilon_r = 2.2$. The structure is modeled by method of moments (MOM) using a mixed-potential integral equation (MPIE) technique [5]. The dielectric constant of the second layer is set to that of air, $\varepsilon_{r1} = 1$. Thickness of the third layer is set to $d_2 = \lambda_2/4$ according to the Type I condition in (1). The results are plotted as a function of d_1/λ_0 for three different values of ε_{r2}. For each set of dielectric constants in Fig 3, the dipole's length is optimized at 31GHz such that the structure is at its resonant, i.e., $\mathrm{Im}(Y_{in})=0$, where Y_{in} is the input admittance seen at the feed-point of the dipole. The results in Fig. 3 show that for this antenna maximum gain occurs for a d_2 value slightly less than that given in (1). It is noteworthy that the presence of the air layer allows an easy mechanical tuning of d_1 in a practical design [6].
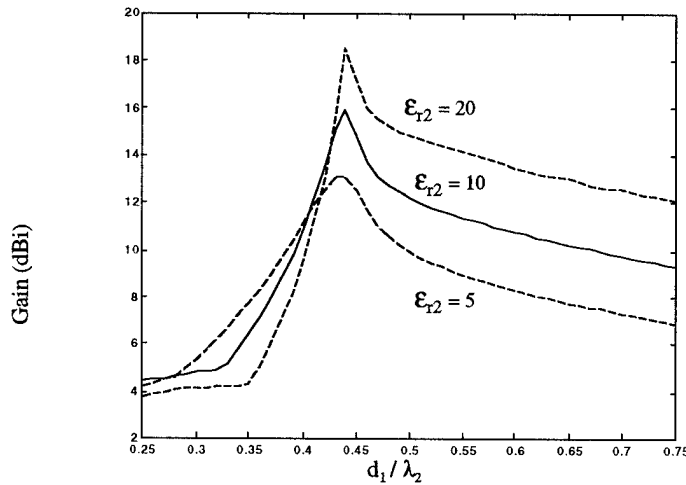

Fig. 3: Gain of the resonant microstrip dipole versus the thickness d_1.

The power coupled into the surface-waves (relative to the radiated power) is shown in Figure 4, where as expected the surface-waves are reduced near the point of maximum gain.
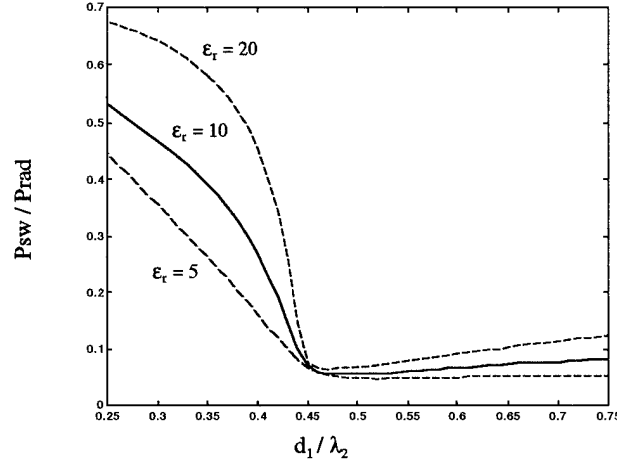

Fig. 4: Power coupled to the surface-waves as a function of d_1.

B. Proximity-fed Microstrip Patch Antenna with a High Permittivity Focusing Layer

The gain enhancement method using a high permittivity focusing layer is inherently narrow band, which makes it difficult to obtain a good much when a microstrip patch is excited via a probe or an edge-fed microstrip line. Proximity-coupled and aperture feeding techniques provide lower level of spurious radiation, wider bandwidth and greater flexibility in impedance matching especially in an array environment. Examples of possible structures utilizing these feeding methods, in conjunction with the multi-layer gain enhancement technique, are shown in Figures 5 and 6 for linear and dual polarization applications, respectively. These configurations can also be easily modified for operation with circular polarization.
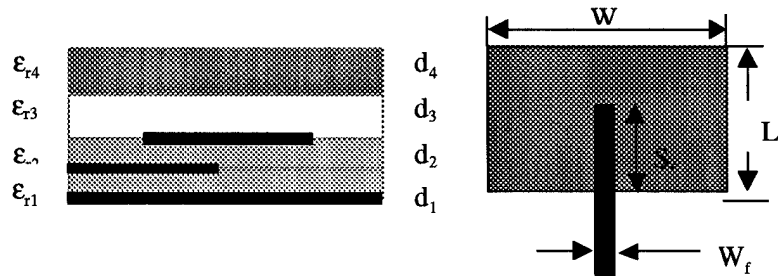
Fig.5: A proximity-fed patch with a high permittivity focusing-layer.
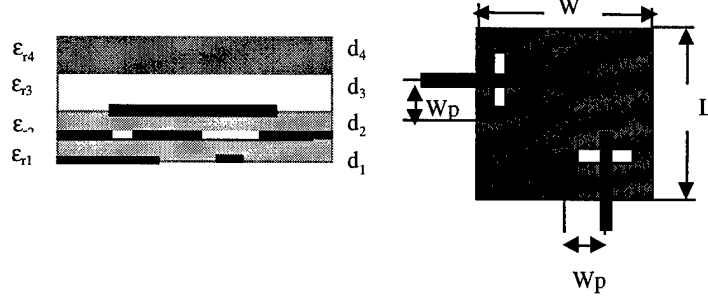
Fig.6: A dual-polarized aperture-fed patch with a high permittivity focusing-layer.

For optimal design of these gain-enhanced antennas we have used an evolutionary optimization in conjunction with method of moments. The multilayer structure is modeled using the MPIE given by [5,8,10]:

$$\left(\frac{-j\eta_0 k_0^2}{2\pi}\right) \sum_{i=1}^{N} \iint_{S_i} \left[\overline{G}_{Mi}^{(p)}(\overline{x},\overline{x}') \ \overline{J}_{si}(\overline{x}') + \frac{1}{k_0^2} \nabla_t \ G_{Ei}^{(p)}(\overline{x},\overline{x}') \ \nabla_t' \cdot \overline{J}_{si}(\overline{x}') \right] dS$$

$$- Z_{sp} \ \overline{J}_{sp} (\overline{x}) = - \overline{E}_{inc}^{(p)} \tag{3}$$

$$\overline{x} = (x,y) \in S \ ; \ p = 1,2,....N$$

where J_{si} and Z_{si} are the electric surface current and the surface impedance of S_i, respectively; E_{inc} is the tangential (incident) electric field impressed at the p-th surface. $G_{Mi}^{(p)}$ and $G_{Ei}^{(p)}$ are the Green's Functions of magnetic and electric types, respectively, evaluated at the p-th layer due to a source at the i-th layer. The MOM solution of (3) with roof-top basis functions is then coupled to a continuous parameter evolutionary programming (EP) algorithm for optimization of the antenna gain [7,8]. As an example the proximity-fed structure in Fig. 5 was optimized for a maximum broadside gain subject to the constraint of a 50 ohm match at 31 GHz with a maximum VSWR of 1.5. The dielectric constants were set fixed to $\varepsilon_{r1} = \varepsilon_{r2} = 2.2$, $\varepsilon_{r3} = 1$ and $\varepsilon_{r4} = 10.2$. The first two layers thickness are $d_1 = d_2 = 0.254$ mm. The remaining parameters, including the length and the widths of the patch, the thickness, d_3 and d_4, of the dielectric layers and the width W_f and the inset length S_f of the microstrip-line, were then optimized using EP, with initial values of d_1 and d_2 chosen according to the condition in (1). The plots of return loss and gain as a function of frequency for the final design are shown in Figure 7. The optimized dimensions are L = 2.57 mm, W = 5.28 mm, $d_3 = 4.38$ mm, $d_4 = 0.593$ mm, $S_f = 0.994$ mm and $W_f = 0.697$ mm. A gain of about 16dBi with VSWR = 1.3 is achieved.
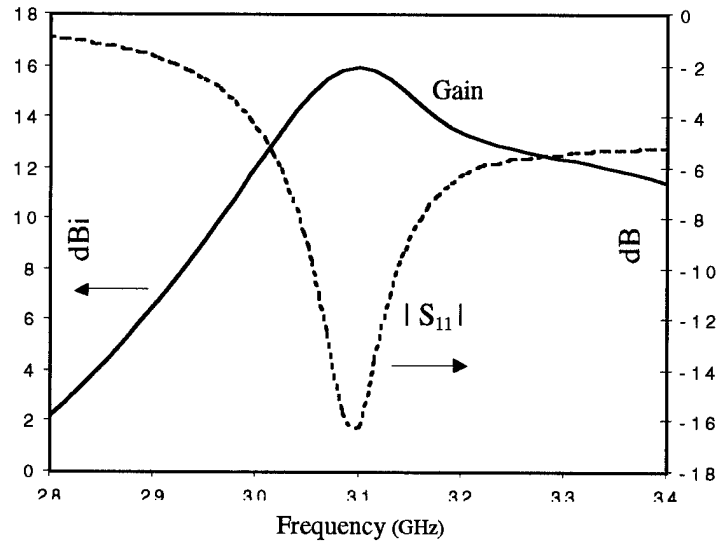
14

Fig. 7: Gain and return loss of the optimized design as a function of frequency

Figure 8 shows the radiation efficiency and the ratio of surface-wave power over radiated power; as seen the power coupled into the surface-waves at the design frequency of 31GHz is minimized, resulting in an efficiency of better than 90%. The dielectric and conductor losses in this case are 0.14 dB and 0.11 dB, respectively. The latter loss is significantly less than that in a standard series or corporate fed microstrip array with a similar gain.
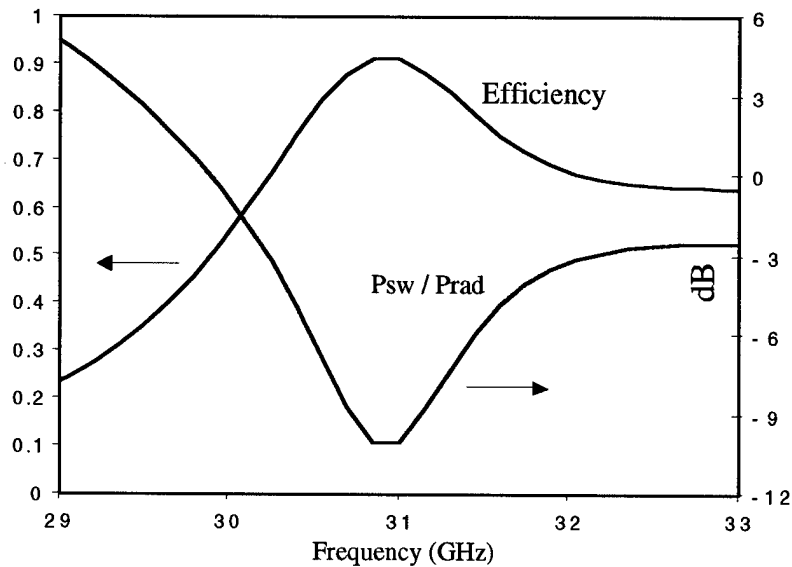
Fig. 8: Radiation efficiency and power coupled to the surface-waves as a function of frequency

The far-field patterns of this antenna are shown in Figure 9. Unlike the standard linear

15

microstrip array, the antenna has nearly equal 3 dB beamwidth in both E- and H-planes; also, the pattern in H-plane ($\phi = 90°$) has a side lobe of about $- 12$dB due to the high permittivity layer.
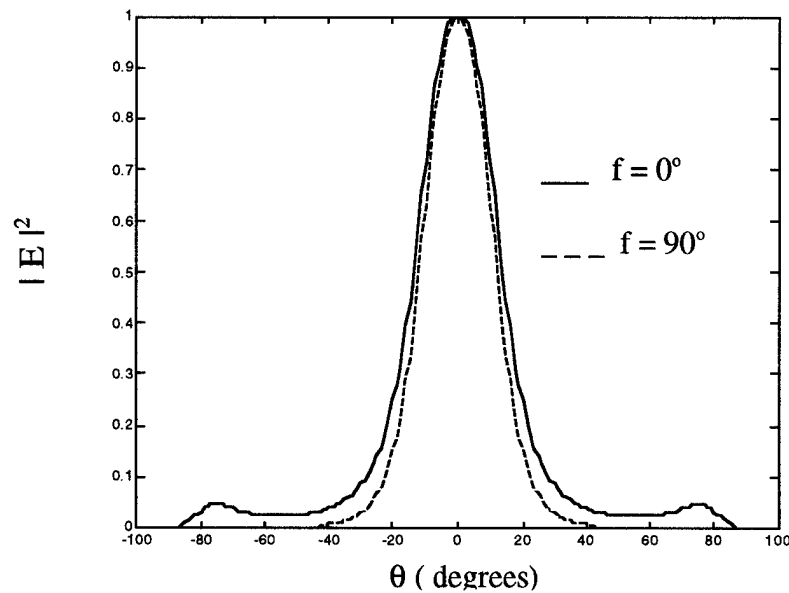
Fig. 9: Power pattern as a function of the elevation angle.

Figure 10 shows the effect of air layer thickness on the gain and VSWR for various values of the focussing layer thickness. As shown high gain together with a good impedance match can be achieved for specific sets of d_3 and d_4 values.
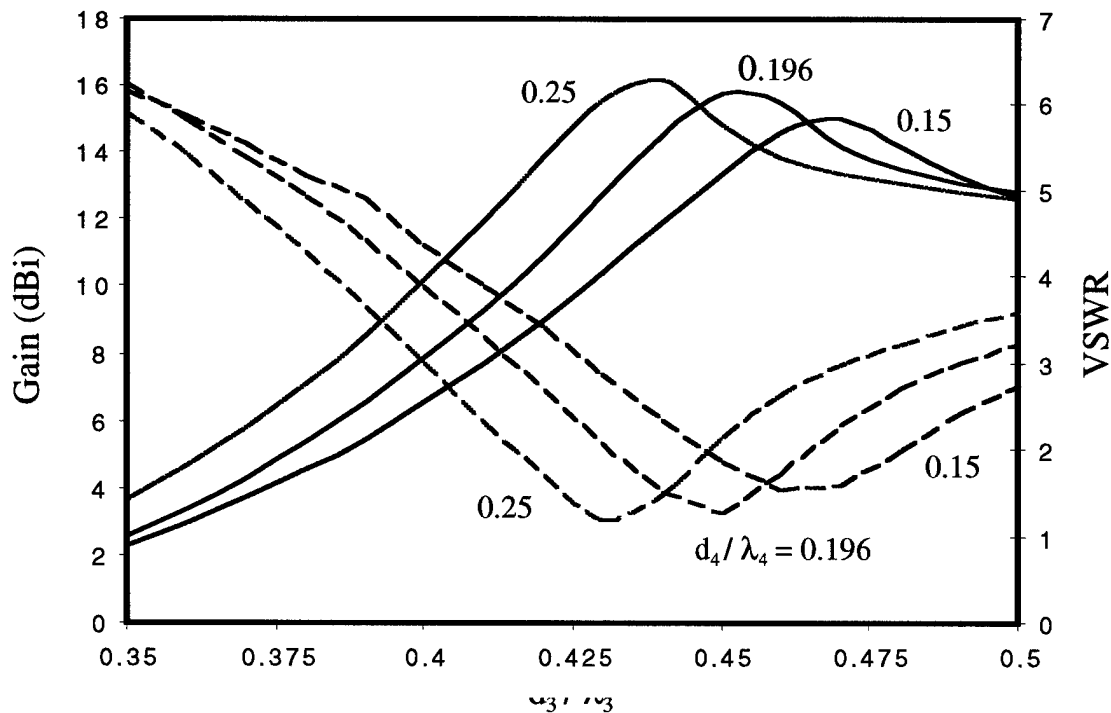

Fig. 10: Gain and VSWR as a function of the air-layer thickness

C. Mutual Coupling Effects

The question naturally arises as how the structures proposed in section II-B perform if they are used as radiating elements in an array environment. Figure 11 shows the H-plane coupling between two four-layer proximity-fed patch elements, separated by an edge-to-edge distance of S. Three cases are included for comparison. In case-1, each antenna element is the structure with $\varepsilon_{r4} = 10.2$, whose radiation characteristics were given in the previous section; case-2 represents a similar structure but optimized for high gain with $\varepsilon_{r4} = 4.5$; case-3 is without any focussing layer, i.e., $\varepsilon_{r4} = 1$. It is clear that the presence of the high permittivity layer significantly increases the coupling between the elements. Since for each high gain element in isolation, the surface wave is minimized (see Fig. 8), one may conclude that the presence of the second element disturbs the overall surface wave structure, resulting in a relatively large value of S_{21}. Figures 12-14 clearly show that as a result of coupling, maximum of the gain shifts to a lower frequency whereas minimum of the P_{sw}/P_r ratio shifts to an upper frequency away from the design frequency of 31 GHz for an element in isolation i.e., $S = \infty$. Nevertheless, for a separation of $S/\lambda_0 = 0.5$ or larger the coupling is negligible. The corresponding patterns for $S/\lambda0 = 0.25$ and 0.5 are shown in Figure 15.
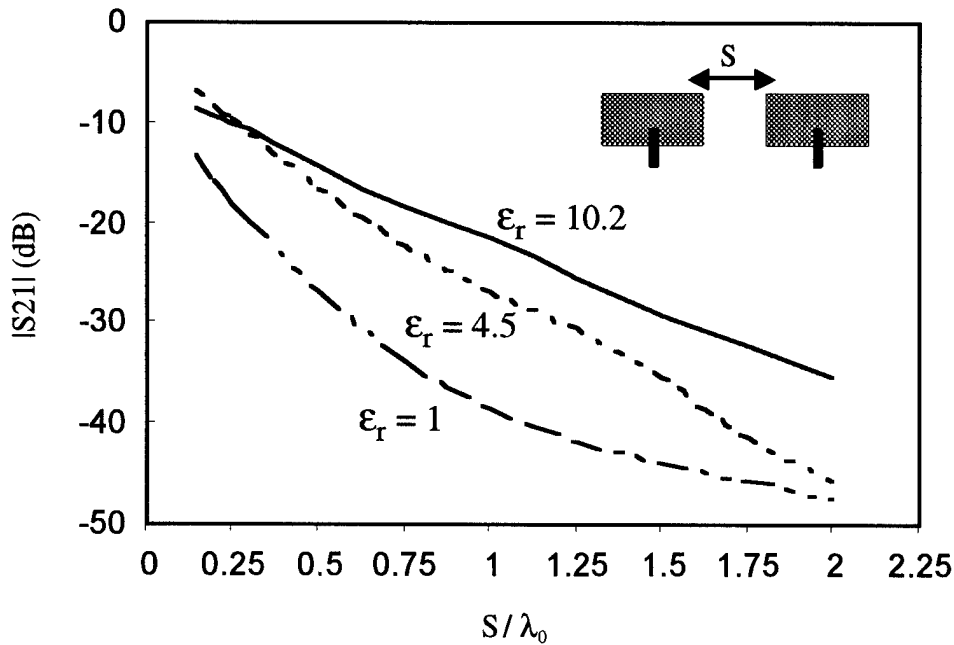

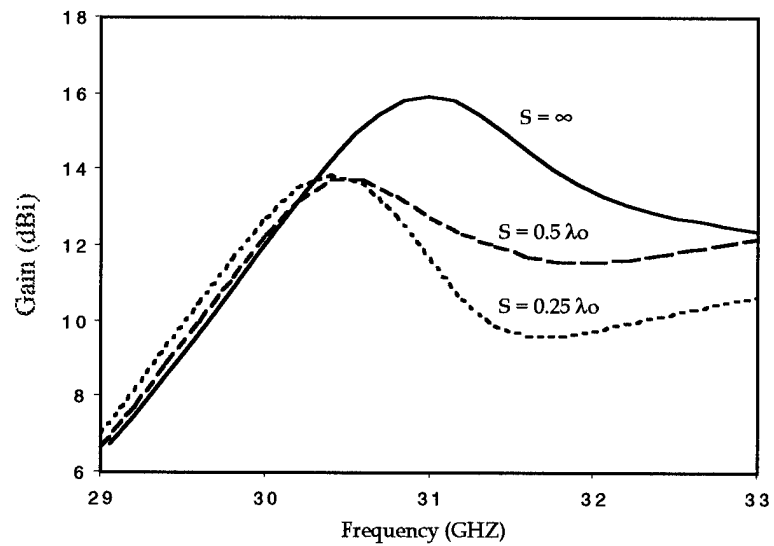Fig. 11: The H-plane coupling between two proximity-fed microstrip patch antennas with a focussing layer.

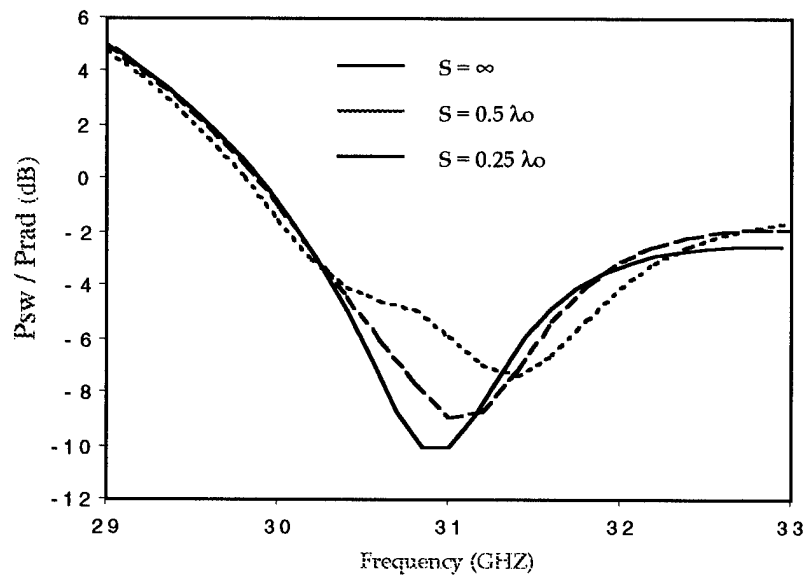Fig. 12: The effect of coupling on the gain as a fuction of frequency.

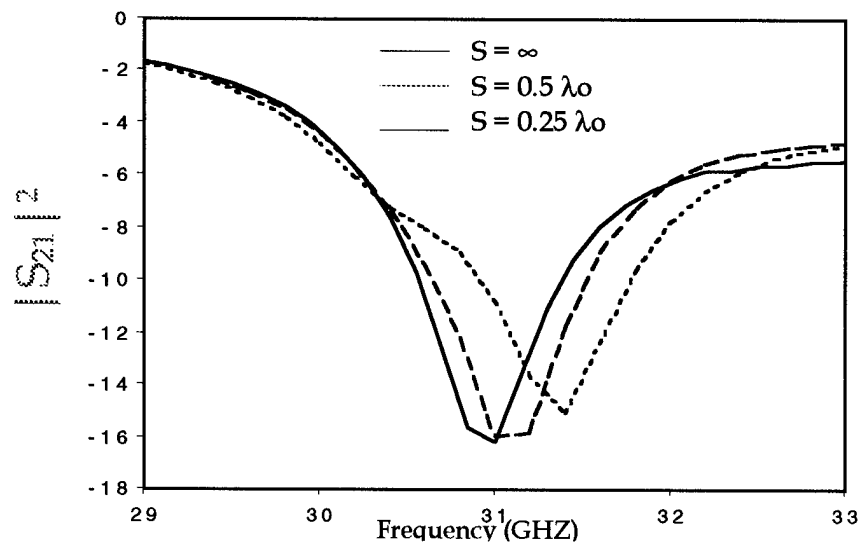Fig. 13: The effect of coupling on the surface-wave power as a function of frequency.

Fig. 14: The effect of coupling on the return-loss as a function of frequency
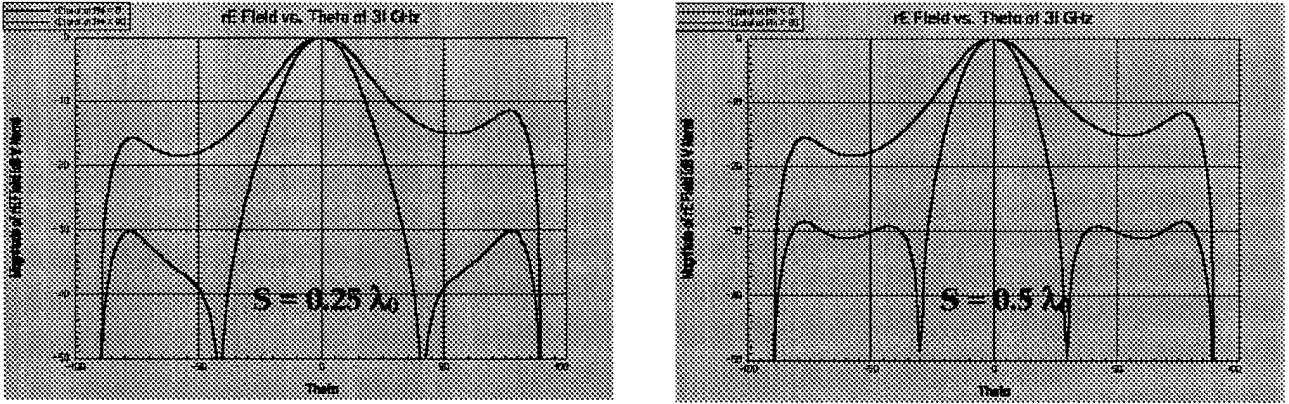
18

Fig. 15: E-plane and H-plane patterns of a two element array of gain-enhanced microstrip antennas

D. Yagi-like Gain Enhancement Method

By introducing parasitic elements at the interface between the dielectric layers, it is possible to obtain high gain without the need for a high permittivity layer [9, 10]. The proximity-fed version of this so-called Yagi-like structure is shown in Figure 16. This structure provides more degrees of freedom than the antenna in Figure 5 in its design optimization.
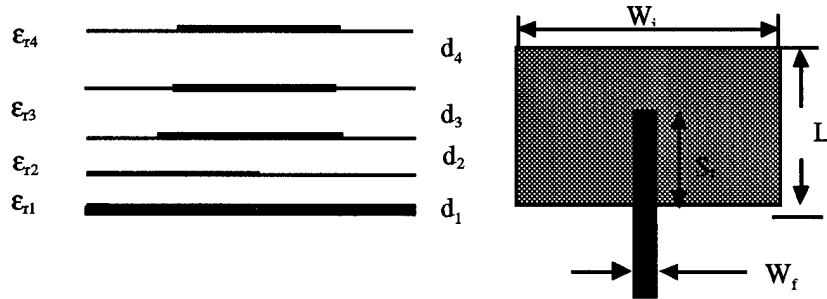

Fig. 16: A Yagi array of proximity-fed stacked microstrip patches.

We have used Evolutionary Programming coupled with the integral equation in (3) to optimize the gain. Two design examples are presented here. In the first example all layers have dielectric constant of 2.2 with $d_1 = d_2 = 0.254$ mm set fixed. All other dimensions in Fig. 16 are optimized subject to the constraint VSWR < 1.5 at 31 GHz. The gain and return loss of the final design are plotted in Fig. 17, where as shown a gain of more than 12 dBi with VSWR = 1.2 is obtained The corresponding radiation efficiency in this case is better than 95%.

19

In the second example the Yagi effect is combined with a high permittivity focussing layer of $\varepsilon_{r4} = 4.5$, and with $\varepsilon_{r1} = \varepsilon_{r2} = 2.2$, $\varepsilon_{r3} = 1$, $d_1 = d_2 = 0.254$ mm. The remaining design parameters are optimized subject to a VSWR < 2 bandwidth constraint of 1 GHz. Figures 18 and 19 show the gain, return loss and patterns for this case. The S_{11} obtained from the commercial code IE3D is also included for comparison. A gain of better than 14 dBi and radiation efficiency of 93% is obtained. The total conductor and dielectric losses is about 0.15 dB. The effect of mutual coupling between two Yagi-like elements, together with the corresponding radiation patterns are shown in Figures 20 and 21. The coupling is negligible for $S \geq 0.5 \lambda_0$.
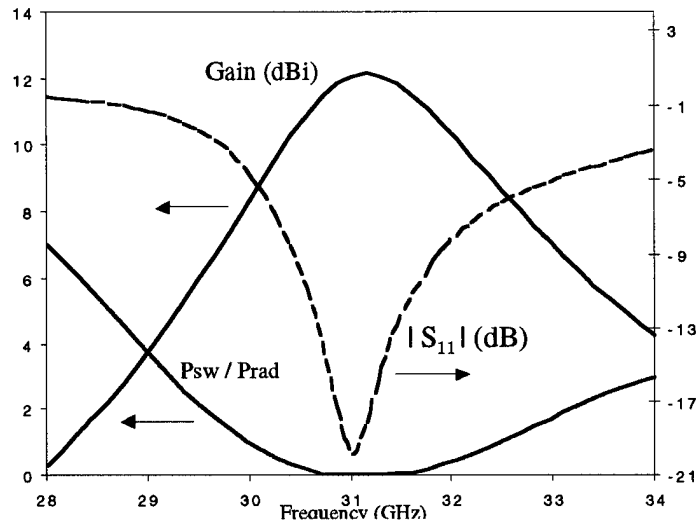

Fig. 17: Gain, surface-wave power and return loss of an optimized design Dielectric constants of all layers are 2.2.
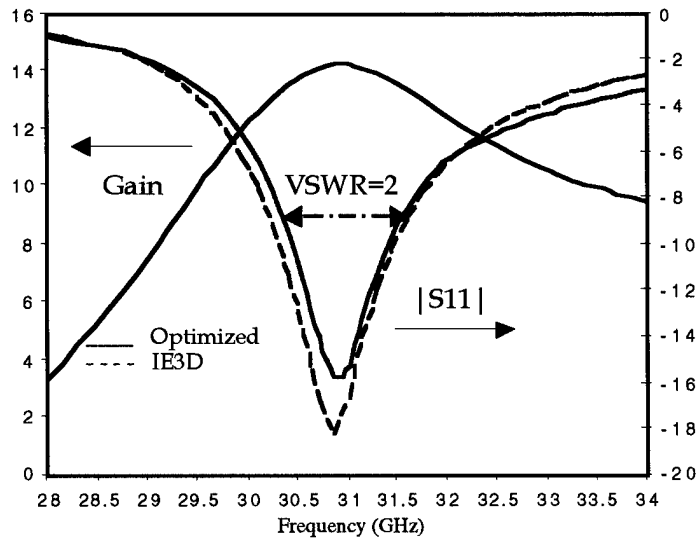

Fig. 18: Gain and return loss of a Yagi-like design, with $\varepsilon_{r4} = 4.5$, as a function of frequency.
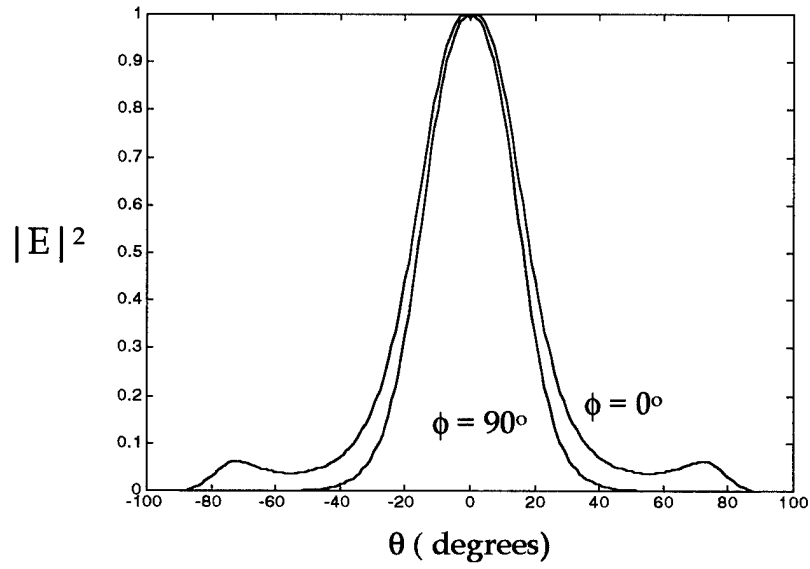
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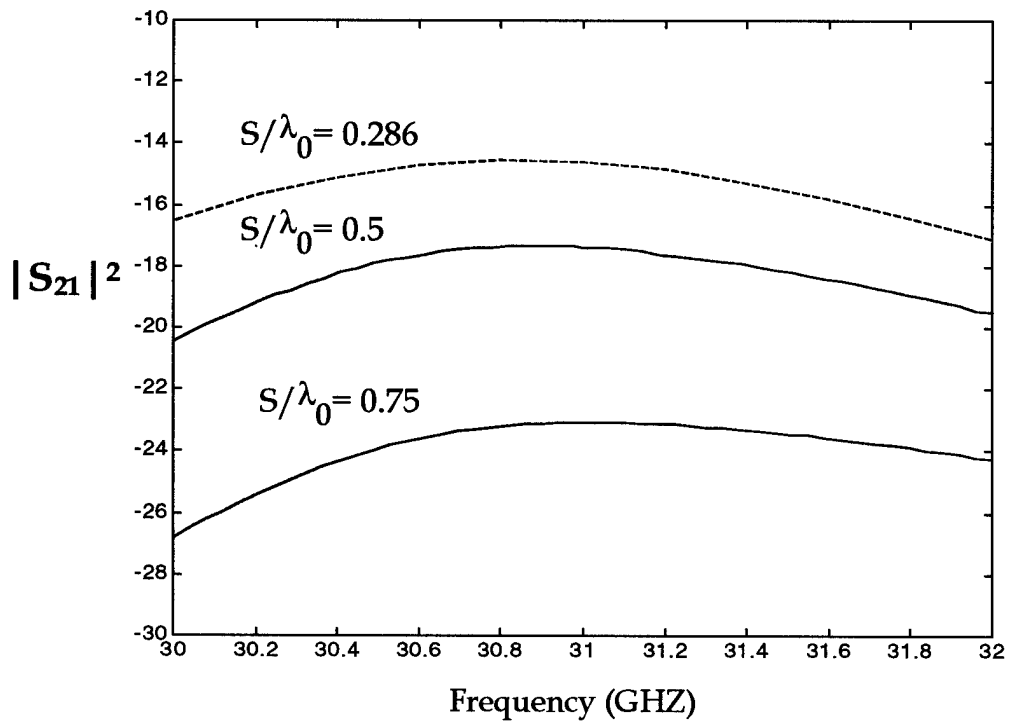Fig. 19: E-plane and H-plane patterns of a Yagi-like microstrip antenna.

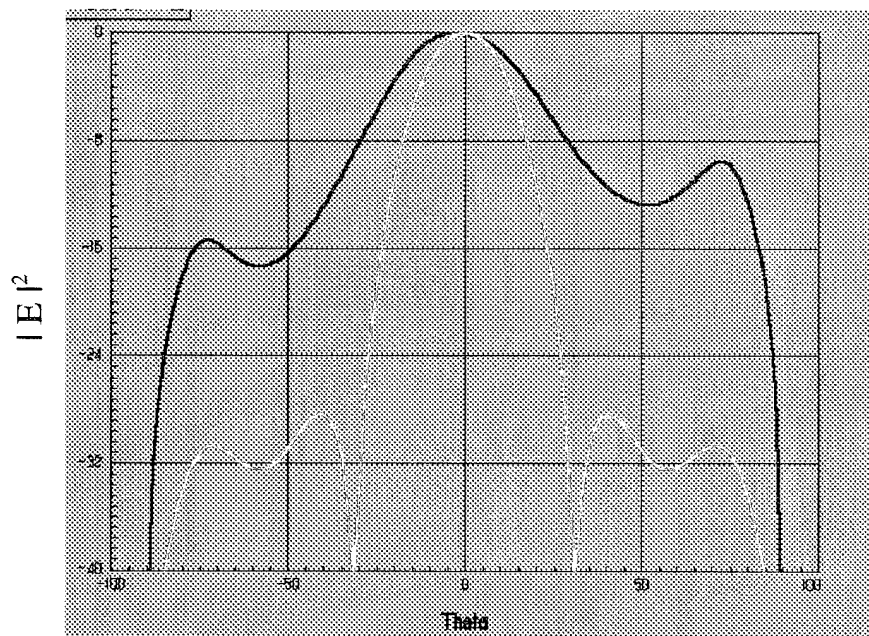Fig. 20: H-plane coupling between two Yagi-like microstrip antenna elements

21

Fig. 21: E-plane and H-plane patterns of a two element array of Yagi-like microstrip antennas, $S = 0.5\lambda_0$

IV. GAIN ENHANCEMENT OF MICROSTRIP ANTENNAS IN A MULTI-LAYERED CYLINDRICAL SURFACE

A. Background

Conventional array methods employing corporate or series feeding techniques have been widely used for design of high gain printed antennas in microwave and millimeter-wave frequencies. Printed arrays, however, suffer from high conductor as well as surface-wave losses that could result in very low antenna efficiency. In addition, these arrays may result in antennas occupying an electrically large real estate, which may not be desirable for some applications. An alternative technique for gain enhancement is based on the resonance condition of a single microstrip radiator in multi-layer geometry with alternating low and high permittivity dielectric layers where the thickness of each layer is about one quarter-wavelength in dielectric [2,3].The Practical configurations of this gain enhancement method, using probe-fed and proximity-fed microstrip antennas, have been reported in [6] and in section II of this report, where a three-layer geometry, consisting of a thin substrate, an air layer of about 0.5 wavelength and a focusing high permittivity layer of about 0.25 wavelength is used.

One of the advantageous features of microstrip antennas is that they are conformable and can be flush-mounted on curved surfaces. The question naturally arises on what effects the curvature might have on the gain enhancement method described above. The purpose of this research is therefore to investigate how antenna directivity and pattern vary with dielectric layers thickness and curvature radius for a gain enhanced microstrip patch on a cylindrical substrate. An exact analysis of such structure requires a full-wave solution using method of moments similar to the one employed in [11]. Assuming an electrically thin substrate, however, the analysis can be made simple by modeling each radiating edge of the rectangular patch with a magnetic line current on a perfectly conducting cylinder loaded with a multi-layered dielectric medium. The formulation is similar to the one previously used in [12] for radiation analysis of a patch antenna on a cylindrical substrate with thick cover layer.

B. Derivation of the Far Fields

Let us first derive expressions for the far-fields radiated from one edge of the patch. Consider Figure 22 showing the cylindrical structure where the inner cylinder of radius $\rho=a$, is

perfectly conducting. The regions a<ρ<b and b<ρ<c constitute the dielectric cover layers whose dielectric constants are $\varepsilon_1 = \varepsilon_o\varepsilon_{r1}$ and $\varepsilon_2 = \varepsilon_o\varepsilon_{r2}$ respectively. Let (b-a) = t_1 and (c-b) = t_2 represent their thickness values. A magnetic line source of strength $\bar{J}_m = \dfrac{p_m}{a}\delta(\phi)\delta(\rho - a)$ is placed along ρ=a, φ=0. Assuming $e^{j\omega t}$ time dependence, we can write the TE$_z$ fields as [12]:

$$E_\phi(\rho,\phi) = -jk_o\eta_o \frac{\partial\Pi_m}{\partial\rho}$$

$$E_\rho(\rho,\phi) = \frac{1}{\rho}\frac{\partial\Pi_m}{\partial\rho}$$

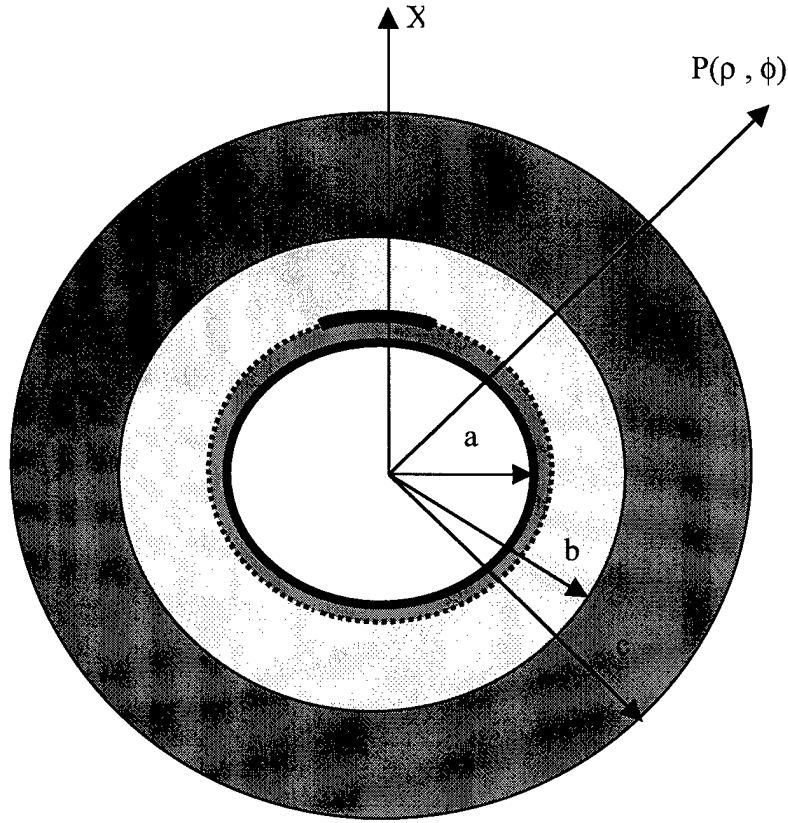$$H_\phi(\rho,\phi) = k^2\Pi_m$$

(4)

Fig.22: Multi-layered cylindrical microstrip patch antenna

where Π_m is the z component of magnetic Hertz vector potential, which satisfies the homogeneous Helmholtz equation for ρ ≠ a . Also k is the wave number that depends on the dielectric constant of the medium in which it is considered. In the two regions a<ρ<b and b<ρ<c

23

we shall denote it by k_1 and k_2 in general, and in the free space region $c<\rho$, $k=k_o$. We might express Π_m as Fourier expansion in different regions as follows:

$$\Pi_m(\rho,\phi) = \begin{cases} \sum_{n=-\infty}^{\infty} \left[\left\{C_n^{(1)} H_n^{(1)}(k_1\rho) + C_n^{(2)} H_n^{(2)}(k_1\rho)\right\}e^{jn\phi}\right] & for\ a < \rho < b \\ \sum_{n=-\infty}^{\infty} \left[\left\{C_n^{(3)} H_n^{(1)}(k_2\rho) + C_n^{(4)} H_n^{(2)}(k_2\rho)\right\}e^{jn\phi}\right] & for\ b < \rho < c \\ \sum_{n=-\infty}^{\infty} \left[\left\{C_n^{(5)} H_n^{(1)}(k_o\rho)\right\}e^{jn\phi}\right] & for\ c < \rho \end{cases} \tag{5}$$

The solution for Π_m for the region $\rho>c$ might be obtained subject to the boundary conditions on the tangential components at $\rho=b$ and $\rho=c$ which are continuous and on the electric field component at $\rho=a$ which is discontinuous by the magnitude of surface magnetic current:

$$\overline{E}^{(1)} \times \hat{\rho}\ \Big|_{\rho=a} = \overline{J}_m \tag{6}$$

Thus we obtain $C_n^{(5)}$ as:

$$C_n^{(5)} = \frac{k_2 k_1^2 W_H(k_2 c) W_H(k_1 b)}{k_o \sum_{i=1}^{8} \zeta_i} \tag{7}$$

where $W_H(x) = \dfrac{4}{j\pi x}$ and

$$\zeta_1 = k_o k_2 H_n^{(1)'}(k_2 c) H_n^{(1)}(k_o c) H_n^{(2)}(k_2 b)\left\{H_n^{(1)'}(k_1 b) H_n^{(2)'}(k_1 a) - H_n^{(2)'}(k_1 b) H_n^{(1)'}(k_1 a)\right\}$$

$$\zeta_2 = k_2^2 H_n^{(1)}(k_o c) H_n^{(1)'}(k_2 c) H_n^{(2)}(k_2 b)\left\{H_n^{(1)'}(k_1 a) H_n^{(2)}(k_1 b) - H_n^{(2)'}(k_1 a) H_n^{(1)}(k_1 b)\right\}$$

$$\zeta_3 = k_o k_1 H_n^{(2)'}(k_2 c) H_n^{(1)}(k_o c) H_n^{(1)'}(k_2 b)\left\{H_n^{(1)}(k_1 b) H_n^{(2)'}(k_1 a) - H_n^{(2)}(k_1 b) H_n^{(1)'}(k_1 a)\right\}$$

$$\zeta_4 = k_o k_2 H_n^{(2)'}(k_2 c) H_n^{(1)}(k_o c) H_n^{(1)}(k_2 b)\left\{H_n^{(2)'}(k_1 b) H_n^{(1)'}(k_1 a) - H_n^{(1)'}(k_1 b) H_n^{(2)'}(k_1 a)\right\} \tag{8}$$

$$\zeta_5 = k_1 k_2 H_n^{(1)'}(k_o c) H_n^{(2)}(k_2 c) H_n^{(1)'}(k_2 b)\left\{H_n^{(2)}(k_1 b) H_n^{(1)'}(k_1 a) - H_n^{(1)}(k_1 b) H_n^{(2)'}(k_1 a)\right\}$$

$$\zeta_6 = k_2^2 H_n^{(1)'}(k_o c) H_n^{(2)}(k_2 c) H_n^{(1)}(k_2 b)\left\{H_n^{(1)'}(k_1 b) H_n^{(2)'}(k_1 a) - H_n^{(2)'}(k_1 b) H_n^{(1)'}(k_1 a)\right\}$$

$$\zeta_7 = k_o k_1 H_n^{(1)'}(k_2 c) H_n^{(1)}(k_o c) H_n^{(2)'}(k_1 a)\left\{H_n^{(1)}(k_1 b) H_n^{(1)'}(k_1 a) - H_n^{(1)}(k_1 b) H_n^{(2)'}(k_1 a)\right\}$$

$$\zeta_8 = k_1 k_2 H_n^{(1)}(k_2 c) H_n^{(1)'}(k_o c) H_n^{(2)'}(k_2 b)\left\{H_n^{(1)}(k_1 b) H_n^{(2)'}(k_1 a) - H_n^{(2)}(k_1 b) H_n^{(1)'}(k_1 a)\right\}$$

By replacing $H_n^{(1)}(k_o\rho)$ with its large argument expansion in (5), one obtains via (4), the expressions for the dominant far-field components E_ϕ and H_z as

$$E_\phi \cong k_o^2\eta_o\sqrt{\frac{2}{\pi k_o\rho}}e^{j(k_o\rho-\pi/4)}\left[C_o^{(5)}+2\sum_{n=1}^{\infty}(j)^{-n}C_n^{(5)}\cos(n\phi)\right] \qquad (9)$$

$$H_z = \frac{1}{\eta_o}E_\phi, \, \eta_o = 120\pi$$

The radiated power per unit length P_r is calculated from the Poynting vector as

$$P_r = \frac{1}{2\eta_o}\int_0^{2\pi}|E_\phi|^2\,\rho d\phi \qquad (10)$$

We can define radiation conductance per unit length of a radiating edge as

$$G_r = \frac{2P_r}{p_m^2} \qquad (11)$$

We shall now consider the case of a microstrip patch shown in Fig.22. Let us assume that the radiating edges of a rectangular microstrip patch of resonant length L_r, are located parallel to the cylinder's axis at $\phi = -\phi_o$ and $\phi = +\phi_o$ respectively. Modeling each edge with a magnetic line source, the total field is obtained as

$$E_\phi = E_{\phi 1} + E_{\phi 2} \qquad (12)$$

where $E_{\phi 1}$ and $E_{\phi 2}$ are now given by the expression in (9) with ϕ replaced by $\phi+\phi_o$ and $\phi-\phi_o$, respectively. Combining the two cosine terms in (9), the far field can be written as

$$E_\phi \cong 2k_o^2\eta_o\sqrt{\frac{2}{\pi k_o\rho}}e^{j(k_o\rho-\pi/4)}\left[C_o^{(5)}+2\sum_{n=1}^{\infty}(j)^{-n}C_n^{(5)}\cos(n\phi_o)\cos(n\phi)\right] \qquad (13)$$

The angle ϕ_o is given by $\phi_o = L_r/2a$, where for $2a/\lambda_o \gg 1$, the resonant length L_r can be approximately determined using the flat ground model,

$$L_r \approx \frac{\lambda_o}{2\sqrt{\varepsilon_{re}}} \qquad (14)$$

In (14) ε_{re} is the effective dielectric constant in the presence of the multi-layered superstrate [13].

In the summations (9) and (13) a large number of terms must be included for an accurate result when the cylinder's radius is large. It was shown in [12] that a minimum of $N > 1.36k_o a$ terms are needed for the convergence of the series in (9) and (13). We have used $N = 2k_o a$ terms in the results presented in this paper.

C. NUMERICAL RESULTS AND DISCUSSION

The overall gain is a complicated function of curvature radius, dielectric constants and thickness values of the superstrate layers. However, assuming an air layer with $\varepsilon_{r1}=1$, for the present geometry, we expect gain enhancement approximately for $t_2/\lambda_2 = 0.25$ and $t_1/\lambda_0 = 0.45$ based on the work in section II for the planar case. Accordingly, we first fix $\varepsilon_r = 10.2$, $t_2/\lambda_2 = .25$ and vary t_1/λ_o in the range $0.2 - 0.55$ and compute directivity for different a/λ_o values. Figure 23 shows the variation in directivity, D_0, in dB as a function of t_1/λ_o over a range $0.2 - 0.5$, for various values of $a/\lambda o$ from 2 to 10. The two dimensional formulation in section III-B, gives the radiated fields only in E-plane. Therefore, in calculation of the directivity we have assumed that the patch has approximately the same field-pattern in both E- and H-plane, consistent with the far-field results presented in section II for the planar case. A similar simulation is carried out for three different permittivity values. Figure.24-a shows variation of D_{0max} in dB as a function of $a/\lambda o$ from 2 to 40, and for various ε_r values of 3.38, 5.12, 6.5, and 10.2. Fig. 24-b shows for what values of t_1/λ_o this D_{max} is obtained when t_2/λ_2 is fixed at 0.25. Next we depict in Fig. 25, variation of D_0 in dB as a function of t_2/λ_2 with t_1/λ_o fixed from Fig. 24-b for maximum D_0 for the choice $a/\lambda o = 10$. The radiation conductance is obtained from (11). In Fig. 26 we show its variation with t_1 and t_2 after normalizing the conductance with respect to that corresponding to $t_1 = t_2 = 0$ case. Figures.26a – 26d depict the four cases for $a/\lambda_o = 20$, 15, 10, and 5 respectively with $\varepsilon_{r2} = 10.2$.

These figures clearly show that the maximum directivity conditions for the thickness of dielectric layers nearly follow those of the planar geometry whenever the radius of the cylinder is about 20 free-space wavelength or larger. For smaller values of radius, however, the curvature has pronounced effects on the directivity and radiation conductance of the antenna, and its effects should be considered for design of multi-layered gain enhanced microstrip patch antennas.
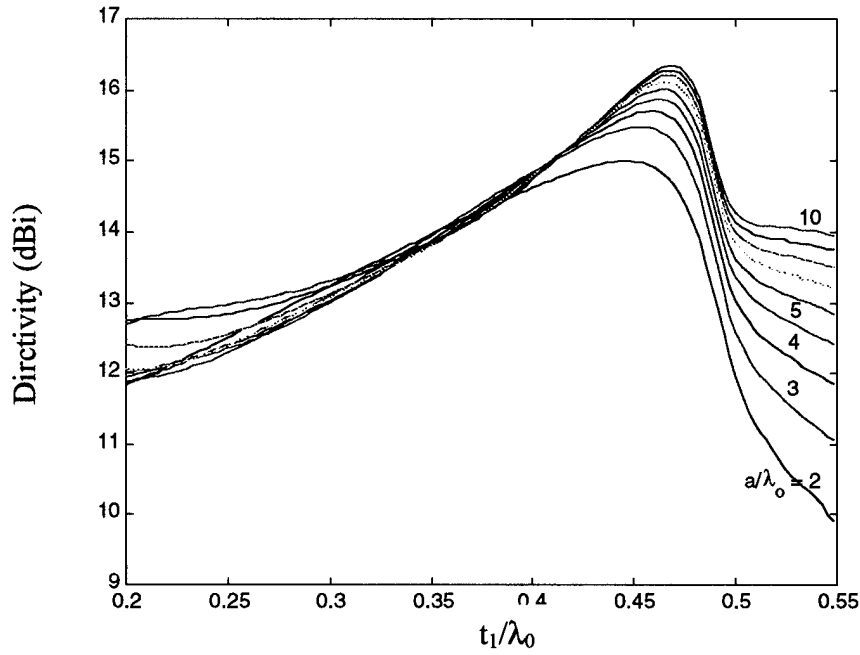
Fig. 23: Variation in directivity as a function of t_1/λ_1 for various values of a/λ_0 from 2 to 10, while $t_1/\lambda_2 = 0.25$.
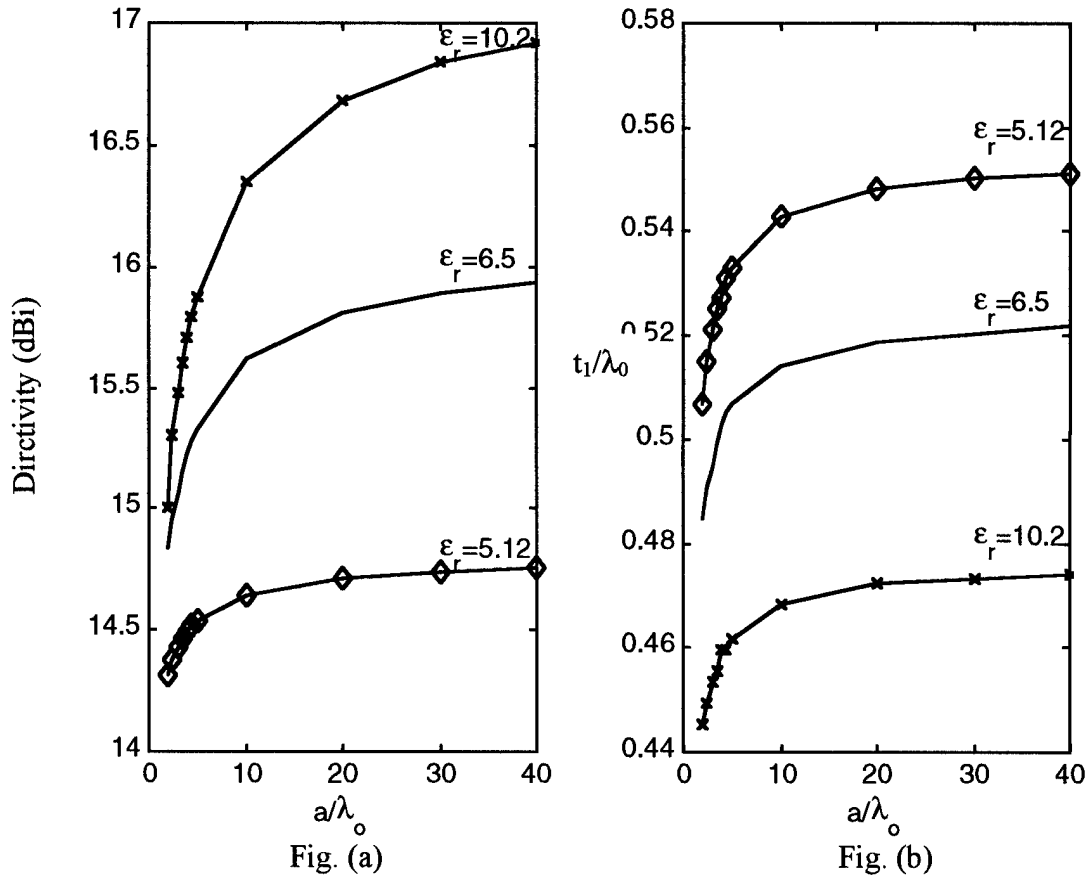

Fig. (a)   Fig. (b)

Fig. 24: (a) Variation in Maximum Directivity as a function of radius in wavelengths and (b) t_1/λ_0 for which maximum directivity is obtained for different dielectrics
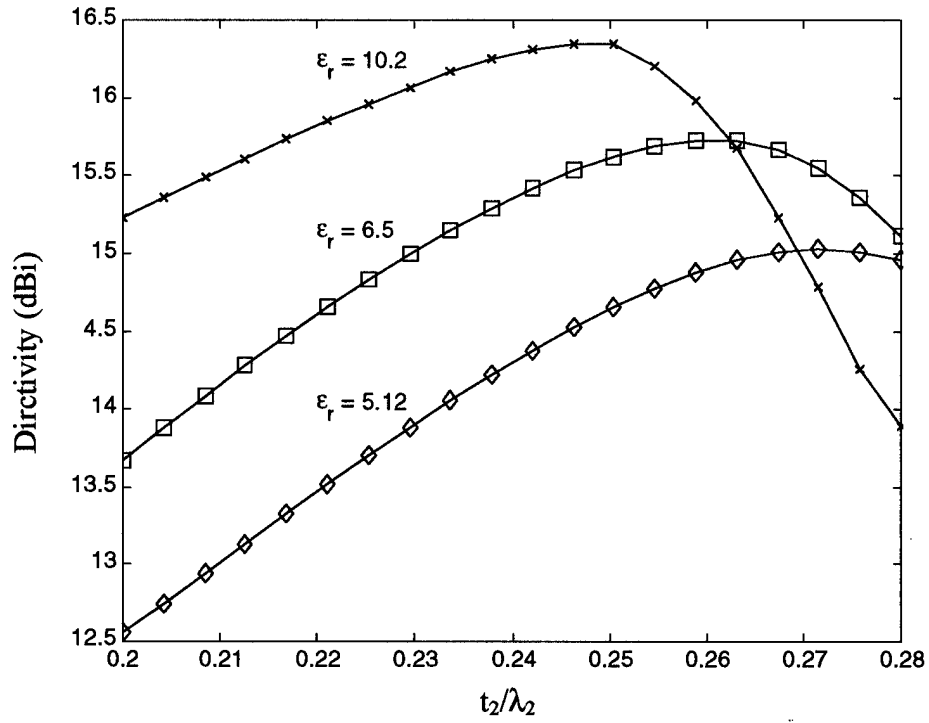
27

Fig. 25: Variation in Directivity with t_2/λ_2 while t_1/λ_o is fixed from the previous simulation shown in Fig. 4b for three values of ε_r.
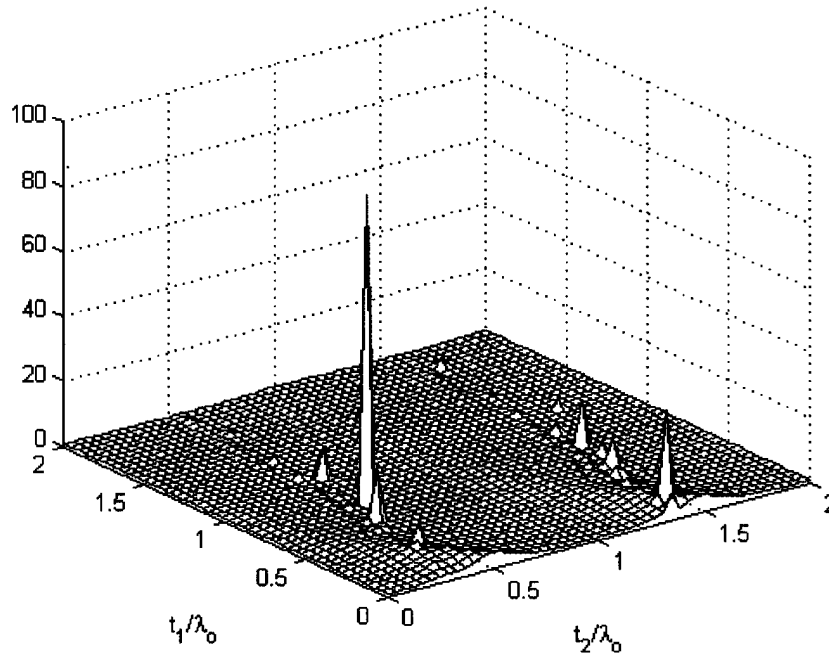
Fig. (a)

Fig. 26: Radiation conductance for different radii in wavelengths. (a) $a/\lambda_o = 20$, (b) $a/\lambda_o = 15$, (c) $a/\lambda_o = 10$, (d) $a/\lambda_o = 5$. It is normalized to the case of $t_1 = t_2 = 0$.
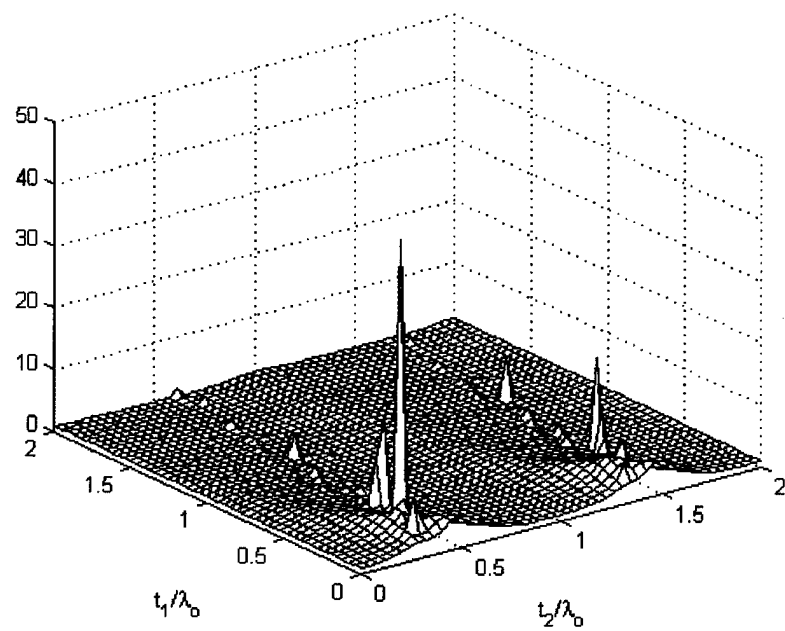
28

Fig. (b)

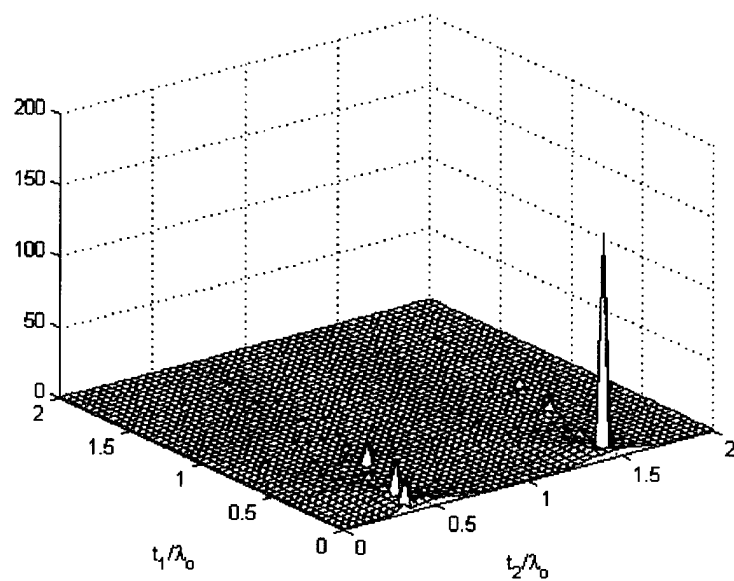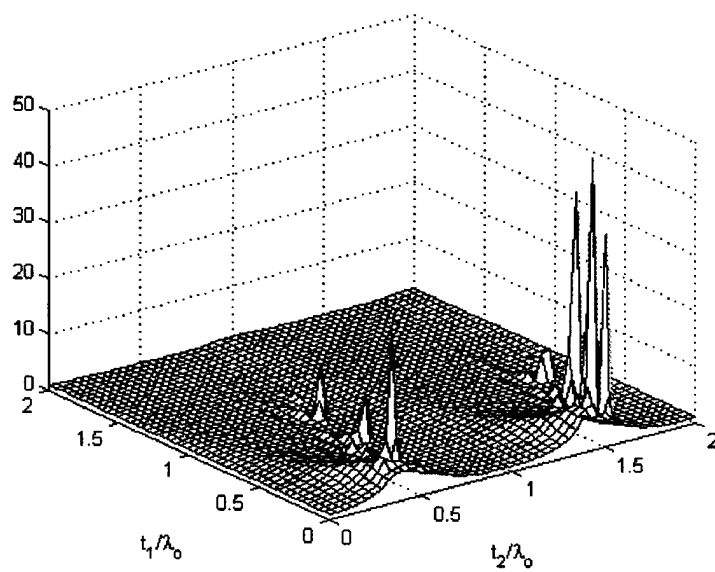Fig. (c)

Fig. (d)

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13. ABSTRACT *(Maximum 200 words)*

The annual reports, previously submitted to the ONR in 1998, 1999 and 2000, have detailed most of our research work on this project. The present report only covers the progress made during the October 2000 - September 2001 interim period. We have made progress in the following area. 1) The previously developed multi-layered gain enhancement technique for printed antennas has been extended to design of proximity-fed and aperture-fed antennas for millimeter-wave applications. Design optimization for the case of a four-layer proximity-coupled patch at a frequency of 31 GHz is presented and its performance in terms of gain, pattern and surface wave efficiency is analyzed. It is shown that conductor losses using this gain enhancement technique, is significantly less than that in a standard microstrip array with a similar gain. The mutual coupling and surface-wave effects in a linear array of such radiating structures are also studied. In addition, an alternative high gain technique, based on the Yagi-like concept that utilizes parasitic elements at the interface between layers has also been investigated in detail for millimeter-wave applications. 2) Effect of curvature on gain enhancement of a multi-layered microstrip antenna is investigated by analyzing the radiated fields of a patch in a multi-layered cylindrical surface. It is shown that curvature effect could significantly change the maximum gain conditions previously reported in the literature for a planar multi-layered microstrip antenna. Development of a FDTD code for accurate characterization of the gain-enhanced printed antennas on a cylindrical surface, is presently in progress

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